DIRECT TORQUE CONTROL OF PERMANENT MAGNET SYNCHRONOUS MOTOR DRIVES WITH CONVENTIONAL AND SVM APPROACH

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By

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Under the Guidance of

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Dedicated to my beloved parents and my all family members.

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CERTIFICATE

This is to certify that the Thesis Report entitled "**DIRECT TORQUE CONTROL OF PERMANENT MAGNET SYNCHRONOUS MOTOR DRIVES WITH CONVENTIONAL AND SVM APPROACH",** submitted by Shree MAHENDRA KUMAR MOHANTY bearing Roll no. 212EE4218 in partial fulfillment of the requirements for the award of Master of Technology in Electrical Engineering with specialization in "Power Electronics and Drives" during session 2012-2014 at National Institute of Technology, Rourkela is an authentic work carried out by him under my supervision and guidance.

To the best of our knowledge, the matter embodied in the thesis has not been submitted to any other University/Institute for the award of any Degree or Diploma.

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ABSTRACT

The vector control of PMSM allows them to be controlled as the same manner as a separately excited DC motor where the torque is directly proportional to the product of armature current and the excitation flux. Here in the case of PMSM the torque control is achieved by controlling the torque component of current and flux component of current independently. This work presents the basic direct torque control scheme and the DTC with space vector modulation technique (DTC-SVM) to improve the performance of Direct Torque Control of Permanent Magnet Synchronous Motor (PMSM) The mathematical model of PMSM with its working principle is presented. Instead of switching table and hysteresis controller, this technique is based on voltage predictive estimation and space vector pulse width modulation. This control scheme improves the torque and flux response and has a constant switching frequency, which reduces the torque and flux ripples with a fast response. The MATLAB/ Simulink Software is used to simulate the scheme and results shows that this DTC-SVPWM scheme has fast dynamic response and very good performance. The DTC-SVM technique improves the performance of basic DTC performances, which features low torque and flux ripples and eliminates the inverter current harmonics witch constant switching frequency.

Chapter I

Introduction

- Permanent Magnet Electric Machines
- Research Background
- Motivation
- Objective
- Dissertation Organization

1. INTRODUCTION:

From the last three decades AC machine drives are becoming more and more popular, especially Induction Motor Drives (IMD) and Permanent Magnet Synchronous Motor (PMSM), but with some special features, the PMSM are usefull to meet advanced necessities such as high power factor, fast dynamic response, and wide range of operating speed like high performance applications, resulting in a gradual increase PMSM the future market in low and mid power applications. Availability of modern Permanent Magnets which have great energy density in the middle of the twentieth century lead to the development of permanent magnet DC machines drives where the rotor is replaced by a permanent magnet. In conventional DC machines, the rotor field is excited through DC exciter which require brushes and commuters, the electromagnets in the rotor replace by the PM, which lead to the compact size of the machine as the electromagnetic rotor winding and the field excitation which requires much more space to accommodate. The PM replaces these windings and external source and hence the machine size is very much reduced. Due to the small size of the permanent magnet motor with high energy density it can be used in many applications where space is a matter of concern. In the absence of brushes and commutators with no external excitation, the PM motors can be used in very sophisticated and hazardous areas with a great safety. Also electrical losses due to field winding of the machine get reduced and also field losses are absent. These improve the thermal characteristics and hence the efficiency. Because of its light weight and small size this motor is more useful in light weight applications like light weight vehicles, hybrid electric vehicles. Like other machines, PM motors also have some disadvantages: the magnets are demagnetized at high temperature, which adds difficulty to the manufacture of permanent magnets and also to the high cost of PM materials.

With the development of soft switching power transistors and silicon controlled rectifiers in the late 1950's, mechanical commutator is replaced by an electronic rectifier in form inverter. The inverter can be controlled by using different switching technologies to get desired operation of the motor. These two developments, i.e. the modern permanent magnets and electronics commutator contributed towards the development of Permanent Magnet motors.

When the electronic commutator replaces the mechanical commutator, the armature needs not be placed on the rotor. The DC machine armature can be on the stator, which enables better cooling; allows greater voltages to be attained. In this case substantial clearance is also offered for insulation purposes. The excitation required on stator is shifted to rotor with the permanent magnet poles.

1.1 PERMANENT MAGNET ELECTRIC MACHINES:

Based on the direction of field flux, Permanent Magnet electric machines are broadly classified into two categories.

i) Axial Field:- Field flux direction is along the axis of the rotor shaft.

ii) Radial Field:- Field flux direction is along the radius of the rotor.

The radial field flux permanent magnet machines are used for low power applications and for high-performance applications, the axial field flux machines used since they have greater power density and higher acceleration capacity.

In many ways, the magnets, can be placed on the rotor. Surface PMs with radial orientation are suitable for high power density synchronous machines, whereas for high speed applications synchronous machines interior PMs are used. Regardless the construction of rotors or placing of

PMs on the rotor the basic working principle of operation for all the machines are same. The important consequence is the dissimilarity between the direct axis and the quadrature axis inductance values. The direct axis is of a rotor is its magnetic axis, which is the principal path of magnetic flux. For high flux density permanent magnets the permeability is comparable to that of air. So we can presume that the magnetic thickness is the addition of the air-gap length by that amount. Direct axis inductance is the inductance when the stator winding is aligned along direct axis or the magnetic axis of the rotor. When the magnets are rotated 90^0 from the aligned position, and the inductance measured in this position is called as the quadrature axis inductance. Based on the mounting of PMs on the rotor the PMSMs are again classified under three categories:

a). Surface Mounted PMSM

The magnets are positioned on the surface of the outer periphery, the motor is called surface mounted PMSM (SPMSM). In this arrangement the magnet in contact with air-gap which results in the greatest air gap flux density. It has lower structural integrity and mechanical robustness. So these machines are favored for low speed applications typically not larger than 3000rpm. Variation of inductances between direct and quadrature axis is very small (less than 10%).

b). Surface Inset PMSM

In this configuration offers a uniform cylindrical surface as the magnets are placed in the grooves of the outer edge of the rotor. It is much more robust in construction, mechanical and it suitable for high speed applications. Here the ratio of quadrature axis to direct axis inductance can be 2 to 2.5 . This configuration is also recognized as inset PMSM.

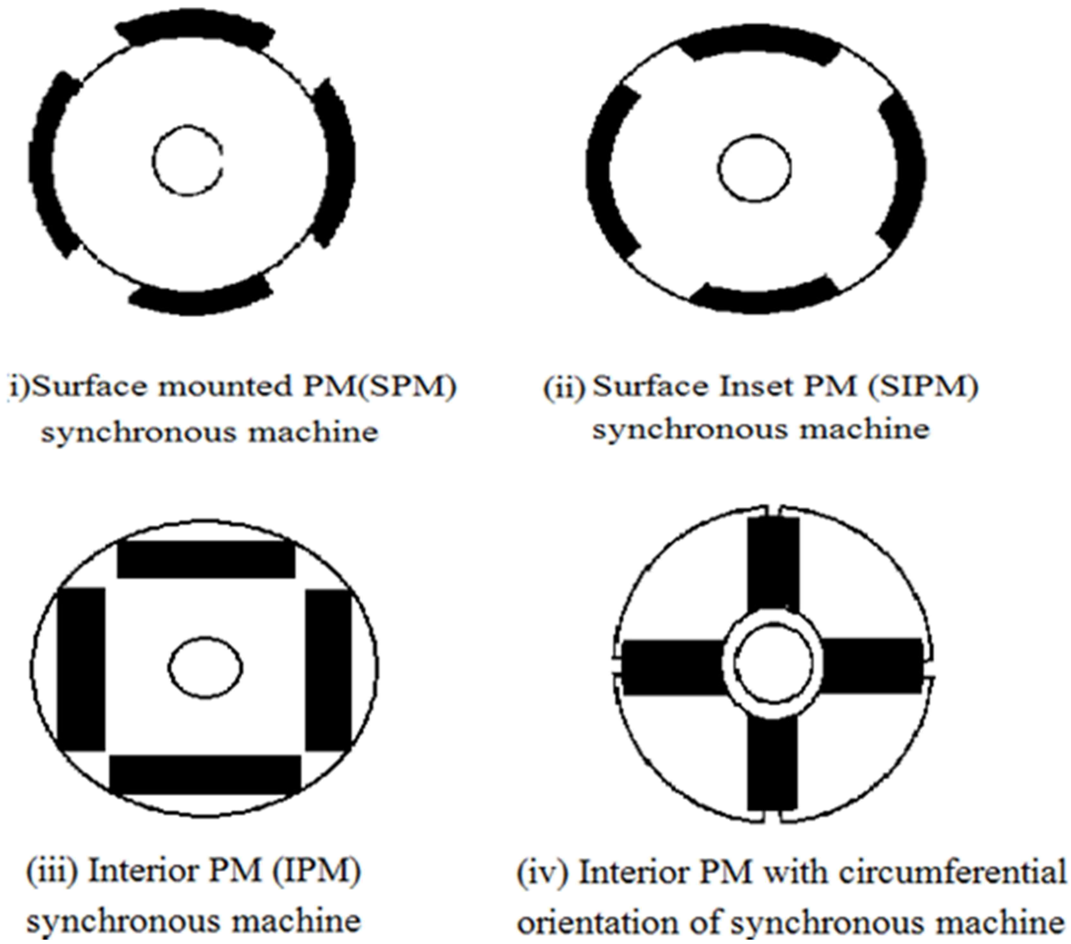


Figure.1.1 (i) Surface PMSM. (ii) Surface inset PMSM. (iii) Interior PMSM. (iv) Interior PMSM with circumferential alignment.

c). Interior PMSM

In the interior PMSM the magnets are buried in the center of the rotor laminations in circumferential and radial directions[5]. This construction is robust, mechanically. The manufacturing of (interior permanent magnets) is more intricate than for the SPM or SIPM rotors. In this case the ratio of the quadrature axis to the direct axis inductance is greater than SIPM rotor but generally does not exceed 3. These are appropriate for high speed applications

From above configurations, inset magnet rotor PMSM has the advantages of both of the SPM and IPM arrangements, i.e. easier construction with mechanical robustness and a high ratio between the quadrature and direct axis inductances.

PM electric machines are also classified into two groups [5]: PMDC machines and PMAC machines. The PMDC machines are similar to the DC commutator machines; the only difference is that the field winding is replaced by the permanent magnets, while in case if PMAC the field is generated by the permanent magnets placed on rotor; the slip rings, the brushes and the commutator does not exist in this machine type. Again the PMAC machines are divided into two types depending on the shape of the back electromotive force (back e.m.f.) induced; trapezoidal or sinusoidal.

The PMAC machines having trapezoidal back e.m.f. are known as brushless DC machines or BLDCMs and the features of this machine are:

- Rectangular dissemination of magnet flux in the air gap
- Rectangular current waveform
- Concerted dissemination stator windings

While permanent magnet synchronous machines or PMSMs have sinusoidal back e.m.f. and the characteristics are:

- Sinusoidal current waveforms
- Sinusoidal dissemination of magnet flux in the air gap

❖ Sinusoidal dissemination of stator conductors.

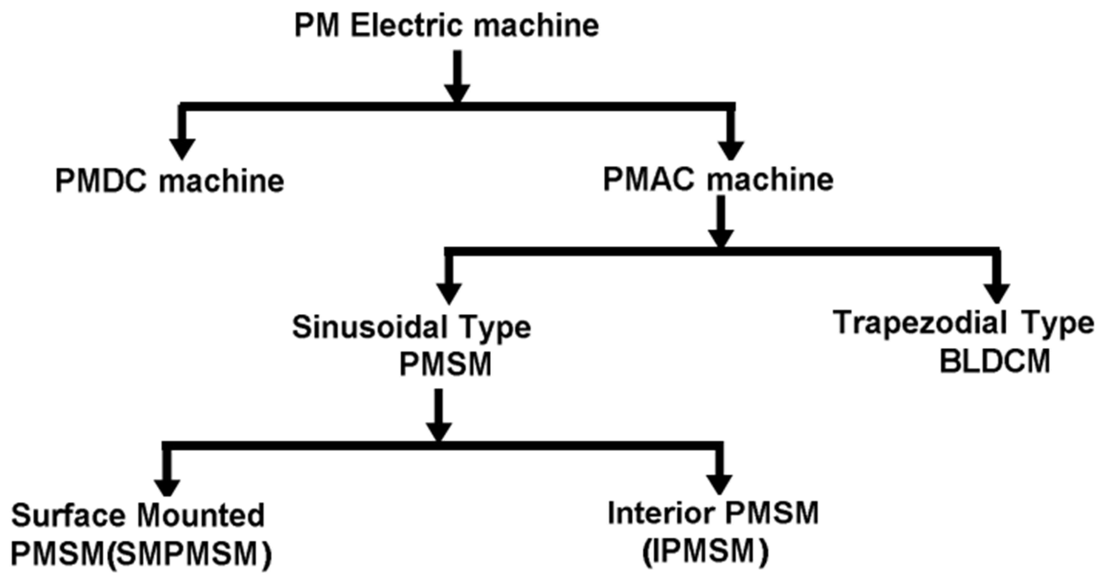


Figure. 1.2 Classification of Permanent Magnet Machines

Around the middle of 1980's, the direct torque control (DTC) topology is first introduced for induction motor drives and the various performances are analyzed. After successful application of the DTC scheme to induction machine drives and getting satisfying results this concept is then applied to the new emerging motor drives, the PMSM drives in the later part of 1980. These PM motors are of low power ratings because of its high cost. Since then the development of different variants of DTC control with better performance is going on.

The direct torque control scheme is a new concept with large prospective for industrial drives[6], especially the PMSM drives. The DTC as the name suggests, is the direct selection of voltage vector for the inverter operation through a lookup table or switching table depending on the difference of reference and actual values of torque and stator flux linkage of the drive system. Hence, we achieve direct control of torque and flux to achieve the desired performance. Here the desired values of flux and torque are estimated by using the voltages and currents of the motor.

Then these estimated values are the actual value of the motor at that instant. So by directly estimating the torque and flux from the currents and comparing them with their reference values we switch the inverter to get the desired operation.

The switching table works on the basis of the position of the rotor also. The voltage vector plane/space is divided into six sectors with an angle difference of 60 degrees. Each vector is spanned between two active vectors. The actual voltage space vector which rotates around the space may be present in any of the sectors. To control that actual voltage the closest voltage vectors are operated. The position of actual voltage vector can be sensed or can be estimated.

There are some problems with this conventional DTC scheme like high torque ripple, current harmonics, drift in flux estimator, not working properly at very low speeds. To get rid of these problems and to achieve better performance different modification from the conventional DTC is applied, discrete space vector modulation is one of them. In space vector modulation technique, the switching period or the sampling time is divided into different sub-periods according to the expected voltage. The voltage vectors operate for a desired time period to obtain the voltage of inverter. The switching frequency is kept constant only the time for voltage vector operation is changed to get the desired voltage. In this case the torque response improved and also we get a high dynamic control.

In this thesis the conventional DTC scheme and the discrete space vector modulation DTC scheme for a surface mounted PMSM is performed in MATLAB simulation and the results have been observed in both of them.

1.2 RESEARCH BACKGROUND:

The application of PMSM motors for high performance applications had made possible by the Vector control techniques where usually only DC motor drives were used[1],[4]-[6]. The vector control of PMSM allows them to be controlled as the same manner as a separately excited DC motor where the torque is directly proportional to the product of armature current and the excitation flux. Here in the case of PMSM the torque control is achieved by controlling the torque component of current and flux component of current independently[1].

Different authors have carried out modelling and simulation of such drives. This section offers a brief review of some of the published work on the PMSM drive system:

Pillay and Krishnan, R. [2] in 1988, presented PM motor drives and classified them into two types viz., PMSM drives and BDCM drives. The PMSM has a sinusoidal back emf and requires sinusoidal stator currents to produce constant torque while the BDCM has a trapezoidal back emf and requires rectangular stator currents to produce constant torque. The dq-axis model of the PMSM were. Equations of the PMSM are derived in the rotor reference frame and the equivalent circuit is presented without dampers.

Further, as an extension of his previous work same author in 1989 [3] presented the application of vector control with complete modelling, simulation, and analysis of PMSM in the rotor reference frame without damper windings. Performance dissimilarities due to the use of hysteresis current controllers and pulse width modulation (PWM) were examined. Specific care was paid to the motor torque pulsations and speed response.

The current-regulated voltage source inverter (VSI) has the advantage of permitting direct torque control by controlling the amplitude of the currents in the machine armature and their phase with respect to the back-emf. A smooth torque generation at low speeds and the system operating limits in the high and extended speed ranges were investigated by Dhaouadi R. and Mohan N. [7] by using ramp, hysteresis and space vector type current controller and performances of these different controllers were also investigated.

Conventional Hysteresis current control technique is widely used because of its fast current control response, inherent peak current limiting capability , and simplicity of implementation. But, a controller with a fixed hysteresis band has the drawback that the modulation frequency varies in a band which results in the generation of non-optimum current ripple at the load. To overcome above mentioned demerits, Bimal. K. Bose [8] suggested an adaptive hysteresis-band current control technique where the band is modulated with the system parameters to preserve the modulation frequency to a constant value. Methodical expressions of the hysteresis band were derived as functions of system parameters.

By this time the direct torque control (DTC) scheme was successfully tested in induction motor drives. So there is an option to examine the effectiveness of this new concept to the PMSM drives. The direct torque control scheme is first proposed by Takahasi and Noguchi [9] in 1985 and then Depenbrock [10][11] in 1988 for the induction motor drives. Takahasi and Noguchi in their work proposed a scheme based on limit cycle control of both the torque and flux by using optimum PWM output voltage; they also proposed a switching table to select the optimum inverter output voltage vectors to achieve most possible fast torque response, low inverter switching frequency, and low harmonic losses. Depenbrock in his work presented the direct self control DSC in which the switches of a three-phase VSI are switched on and off directly thru three Schmitt triggers, by comparing the time integrals of line-to-line voltages to a reference value of desired flux.

The DTC scheme then first applied to PMSM drives by M.F. Rahman and L. Zhong [12] - [14]. They showed that change of electromagnetic torque in a PM motor is proportionate to the change of the angle between the stator and rotor flux linkages, and, therefore, the fast torque response can be obtained by adjusting the rotating speed of the stator flux linkage as fast as possible. They had also shown that the zero voltage vectors should not be used, and stator flux linkage should be kept moving with respect to the rotor flux linkage all the time.

The advantage of DTC are low complexity and it requires only one parameter, the stator resistance. No PWM method is required. All the estimation or formulation are done in the stationary reference frame, no explicit knowledge about the rotor position is required [12]. But the rotor position at the start up must be known. The system possesses good dynamic behavior, but it gives poor performance in steady state with high levels of ripples in torque and flux linkages and in stator currents[15]-[17]. This happens due to the crude voltage vector selection for inverter switches[18].

To overcome these disadvantages, different control strategies are proposed[15]-[23], space vector modulation (SVM) technique is one of the methods[22]. This is a pulse width modulation technique which synthesizes any of the voltage vectors lying inside the sector covered by the six VSI voltage vectors. In this DTC-SVM scheme the hysteresis band controllers are substituted by an estimator which compensates the torque and flux errors, and also instead of switching table a voltage modulator is used.

In 2002, D. Swierczynski and M. P. Kazmierkowk [22] in their paper first presented the DTC-SVM technology for the PMSM drives. They able to improve the performance of the drive by applying SVM concept. The results show that the DTC-SVM not only reduces the ripples, but also gives constant switching operation. In 2004, Jin Mengjia, Shi Cenwei, Qiu Jianqi, Lin Ruiguang[24], in their work presented a stator flux estimation scheme for direct torque controlled (DTC) surface mounted permanent magnet synchronous motor (PMSM) drives, which achieves better low speed performance. Here, in the full speed region the torque ripple and current distortion keep low. Lin Wiejie [25] investigated a direct torque control (DTC) strategy based on observing stator flux linkage for permanent magnet synchronous motor (PMSM) with SVM.

1.3 MOTIVATION:

Comprising with above mentioned many special features and characteristics of PMSM, it has been found very interesting subject matter for the present researchers. PMSM drive is largely maintenance free, which ensures the most efficient operation and it can be operated at improved power factor which can help in improving the overall system power factor and eliminating or reducing utility power factor penalties. From the research over PMSM until now it shows that, in the future market PMSM drives could become an emerging competitor for the Induction motor drive in servo application and many industrial applications. So now there is a great challenge to improve the performance with accurate speed tracking and smooth torque output, minimizing its ripple during transient as well as steady state condition such that it can meet the expectation of future market demand.

So looking out with such a motive, here a speed controller having superior performance for speed tracking has been designed as outer loop and a current controller, which can provide smooth, ripple less torque response has also been designed as inner loop for closed loop operation of the drive. Due to the cost, modelling and simulation is generally cast-off in designing PM drives compared to building system prototypes. After selecting the component model simulation process can start to calculate the steady state and dynamic performance. This practice cuts time, cost of constructing prototypes. So, Simulations helps in developing new systems, by reducing costs and which is done here in MATLAB/Simulink platform.

1.4 OBJECTIVE:

The foremost objective of this research is to improve the performance of an SPMSM drive system by achieving more precise speed tracking and smooth torque response by implementing the Conventional Diect Torque Control Scheme and a Discrete Space Vector Modulation DTC scheme respectively by employing their superior performance.

The overall objectives to be achieved in this study are:

- To design the equivalent d-q model of SPMSM for its vector control analysis and closed loop operation of the drive system.
- Analysis and implementation of conventional DTC scheme in steady state and transient condition (step change in load and speed) in MATLAB/Simulink environment.
- Analysis and implementation of Discrete SVPWM-DTC control scheme to PMSM during steady state and transient condition in MATLAB/Simulink environment.

- Comparison of system performance using conventional DTC and SVM-DTC schemes in MATLAB/Simulink environment to compare their performances so as to consider better controller for our system applications.

1.5 DISSERTATION ORGANIZATION:

The dissertation is organized as follows:

Chapter 1 presents the background for this thesis research, motivation and the research objectives along with a comprehensive literature review in related areas is also given.

Chapter 2 includes the mathematical modelling of surface permanent-magnet synchronous machines in the rotor reference frame. Moreover, basic vector control operation principles of PM synchronous machines are briefly discussed.

Chapter II

Overview and Dynamic Modelling of Permanent Magnet Synchronous Motor Drive System and Common Control Strategies

- Mathematical Model of PMSM in abc-frame
- Dynamic d-q Modelling
- Park Transformation
- Equivalent circuit of PMSM
- Some common control schemes for PM motors
  - Scalar Control
  - Vector Control
    - Field Oriented Control
    - Direct Torque Control
    - DTC-Space Vector Modulation control

Overview and Dynamic Modelling of Permanent Magnet Synchronous Motor Drive System and Common Control Strategies:

Here the motor model is derived which used for simulations. This model emphases on fundamental frequency component. Stator windings are considered distributed sinusoidally, the magnetic circuits are presumed linear, and the magnetic fields are presumed identical in the rotor axis, so there is no effect of end fringing.

2.1. Mathematical Model of PMSM in abc-frame:

The PMSM is a three phase alternating current machine as shown in figure 2.1, where the stator windings are portrayed as single coils together with their magnetic axes.

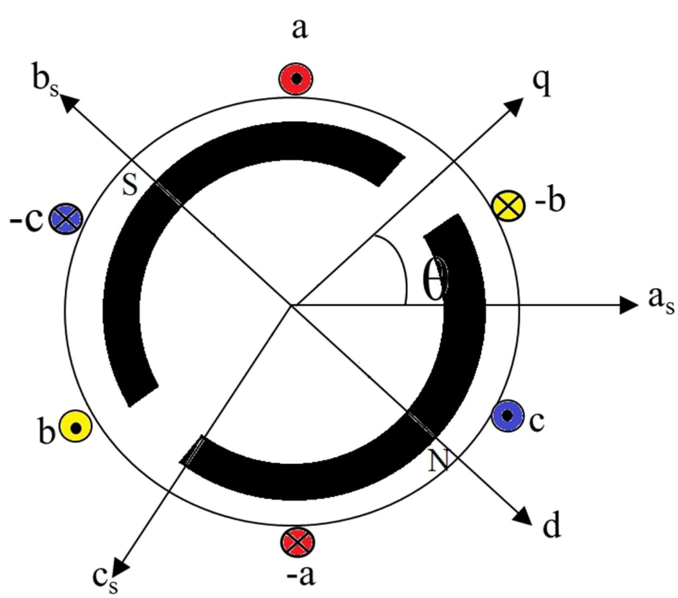


Figure.2.1, A simplified outline showing a SPMSM motor's stator windings, rotor magnet and their magnetic axes.

Almost all the magnetic fields are concerted to air-gap as the material which used for stator and rotor are made of low reluctance than the air-gap between them, and therefore. the fields can assumed constant through the air-gap, as the air-gap is small compared to the rotor radius.

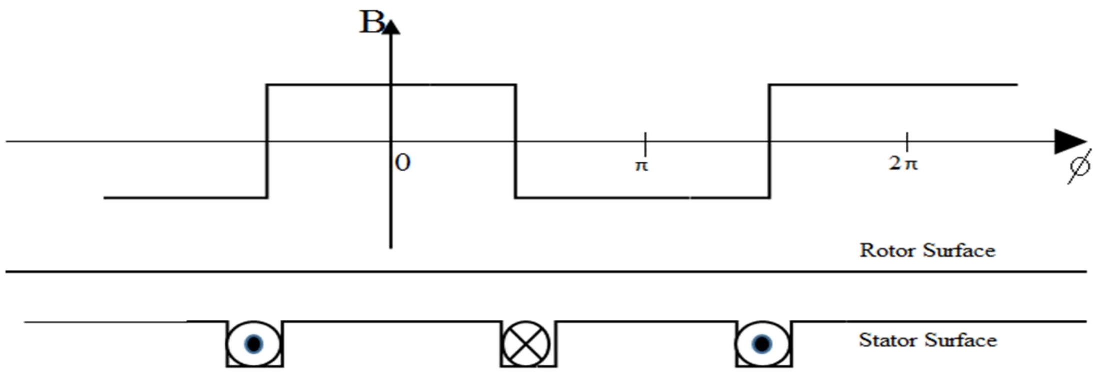


Figure. 2.2 Magnetic-field of a concentrated winding.

The resultant field more look like a sine-wave when stator winding is spread over several slots.

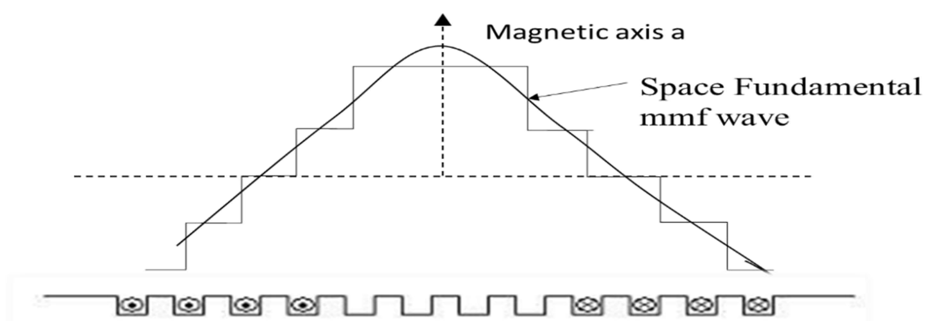


Figure 2.3, Magnetic field of a distributed winding

PMSM Machine Equations

The electrical angle/speed is different from the mechanical angle/speed in a multimagnet machine(i.e. the no. of pole pairs greater then 1).

$$\omega_r = \frac{P}{2}\omega_m, \tag{2.1}$$

Where ω_r is the rotor speed measured in electrical degrees, ω_m in mechanical degrees and P is the number of poles.

In each stator winding the voltage, V, is the sum of the resistive voltage drop and the voltage induced from the time dependent flux linkage, ${d\psi}/_{dt}$,

$$V_a = R_a i_a + \frac{d\psi_a}{dt}$$

$$V_b = R_b i_b + \frac{d\psi_b}{dt} \qquad (2.2)$$

$$V_c = R_c i_c + \frac{d\psi_c}{dt}$$

The resistances in all the windings are equal, because all the stator windings are wound with equal number of turns, i.e.

$R_a = R_b = R_c = R_s$

In matrix form,

$$\boldsymbol{V}_{abc} = \boldsymbol{R}_s \boldsymbol{i}_{abc} + \frac{d}{dt}\psi_{abc} = \begin{bmatrix} R_s & 0 & 0 \\ 0 & R_s & 0 \\ 0 & 0 & R_s \end{bmatrix} \begin{bmatrix} i_a \\ i_b \\ i_c \end{bmatrix} + \frac{d}{dt} \begin{bmatrix} \psi_a \\ \psi_b \\ \psi_c \end{bmatrix} \qquad (2.3)$$

And

$$\boldsymbol{\psi}_{abc} = \boldsymbol{L}_s \boldsymbol{i}_{abc} + \boldsymbol{\psi}_m \qquad (2.4)$$

Where $\boldsymbol{L}_s$ is the stator inductance and $\boldsymbol{\psi}_m$ is the flux linkage in stator windings due to the Permanent Magnet.

2.2.Dynamic d-q Modelling

The mathematical model for the vector control of the PMSM can be derived from its dynamic d-q model with the equation of damper winding and field current dynamics removed. The synchronously rotating rotor reference frame is chosen so the stator winding quantities are transformed to the synchronously rotating reference frame that is revolving at rotor speed.

The model of PMSM is developed on rotor reference frame using the following assumptions:

- Saturation is ignored.
- Sinusoidal induced EMF.
- Negligible Core losses.
- There are no field current dynamics.

It is also be assumed that rotor flux is constant at a given operating point and along the d-axis zero flux linkage along the q-axis[1].

Since the position of the rotor magnets determine the instantaneous induced emf and then the stator currents and torque of the machine the rotor reference frame is selected independent of the stator parameters. Here the equivalent d and q axis stator windings are transformed to the reference frames that are revolving at rotor speed. So there is zero speed differential between the stator and rotor magnetic fields. The stator d and q axis windings have a fixed phase relationship with the rotor magnet axis that is the d axis. The stator equations of the induction machine in rotor reference frame using flux linkages are in use to derive the model of the SPMSM as shown in Fig.2.1.

So an SPM machine is described by the following set of general equations:

Stator voltage equations are:

$$V_d = R_s i_d - \omega_r \psi_q + \frac{d\psi_d}{dt} \tag{2.5}$$

$$V_q = R_s i_q + \omega_r \psi_d + \frac{d\psi_q}{dt} \tag{2.6}$$

Flux linkages are given by

$$\psi_d = L_d i_d + \psi_{PM} \tag{2.7}$$

$$\psi_q = L_q i_q \tag{2.8}$$

Replacing equations (2.7) and (2.8) into (2.5) and (2.6), we get

$$V_d = R_s i_d - \omega_r L_q i_q + \frac{d}{dt}(L_d i_d + \psi_{PM}) \tag{2.9}$$

$$V_q = R_s i_q + \omega_r (L_d i_d + \psi_{PM}) + \frac{d(L_q i_q)}{dt} \tag{2.10}$$

Arranging equations (2.20) and (2.21) in matrix form

$$\begin{bmatrix} V_d \\ V_q \end{bmatrix} = \begin{bmatrix} R_s + \frac{dL_d}{dt} & -\omega_r L_q \\ \omega_r L_d & R_s + \frac{dL_q}{dt} \end{bmatrix} \begin{bmatrix} i_d \\ i_q \end{bmatrix} + \begin{bmatrix} \frac{d\psi_{PM}}{dt} \\ \omega_r \psi_{PM} \end{bmatrix} \tag{2.11}$$

The electromagnetic torque developed by motor is given by

$$T_e = \frac{3}{2}\left(\frac{P}{2}\right)(\psi_d i_q - \psi_q i_d) \tag{2.12}$$

$$T_e = \frac{3}{4} P[\psi_{PM} i_q + (L_q - L_d) i_d i_q \tag{2.13}$$

The mechanical torque equation is

$$T_e = T_L + B\omega_m + J\frac{d\omega_m}{dt} \tag{2.14}$$

Solving for rotor mechanical speed from (2.25), we get

$$\omega_m = \int \left(\frac{T_e - T_L - B\omega_m}{J}\right) dt \tag{2.15}$$

And rotor electrical speed is $\omega_r = \omega_m \left(\frac{P}{2}\right)$ (2.16)

2.3.Park Transformation:

For the study of motor during transient and steady state conditions the dynamic d-q modelling is used. Here the three phase voltages and currents are converted from abc-variable to dq0 variables by using Parks transformation [5]. The following equations are obtained after after conversion of the phase voltages variables V_{abc} to V_{dq0} variables in rotor reference frame:

$$\begin{bmatrix} V_q \\ V_d \\ V_0 \end{bmatrix} = \frac{2}{3}\begin{bmatrix} \cos\theta_r & \cos(\theta_r - 120) & \cos(\theta_r + 120) \\ \sin\theta_r & \sin(\theta_r - 120) & \sin(\theta_r + 120) \\ 1/2 & 1/2 & 1/2 \end{bmatrix}\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} \tag{2.17}$$

In contrast, V_{dq0} can be converted to V_{abc} as:

$$\begin{bmatrix} V_a \\ V_b \\ V_c \end{bmatrix} = \frac{2}{3}\begin{bmatrix} \cos\theta_r & \sin\theta_r & 1 \\ \cos(\theta_r - 120) & \sin(\theta_r - 120) & 1 \\ \cos(\theta_r + 120) & \sin(\theta_r + 120) & 1 \end{bmatrix}\begin{bmatrix} V_q \\ V_d \\ V_0 \end{bmatrix} \tag{2.18}$$

2.4.**Equivalent circuit of PMSM:**

Equivalent circuits of the motors are used for study and simulation of motors. Considering rotor d-axis flux from the permanent magnets is denoted by a constant current source as described in the equation $\psi_{PM} = L_{dm} i_f$, following figures can be obtained from [1] shown as fig 2.4 and fig.2.5.

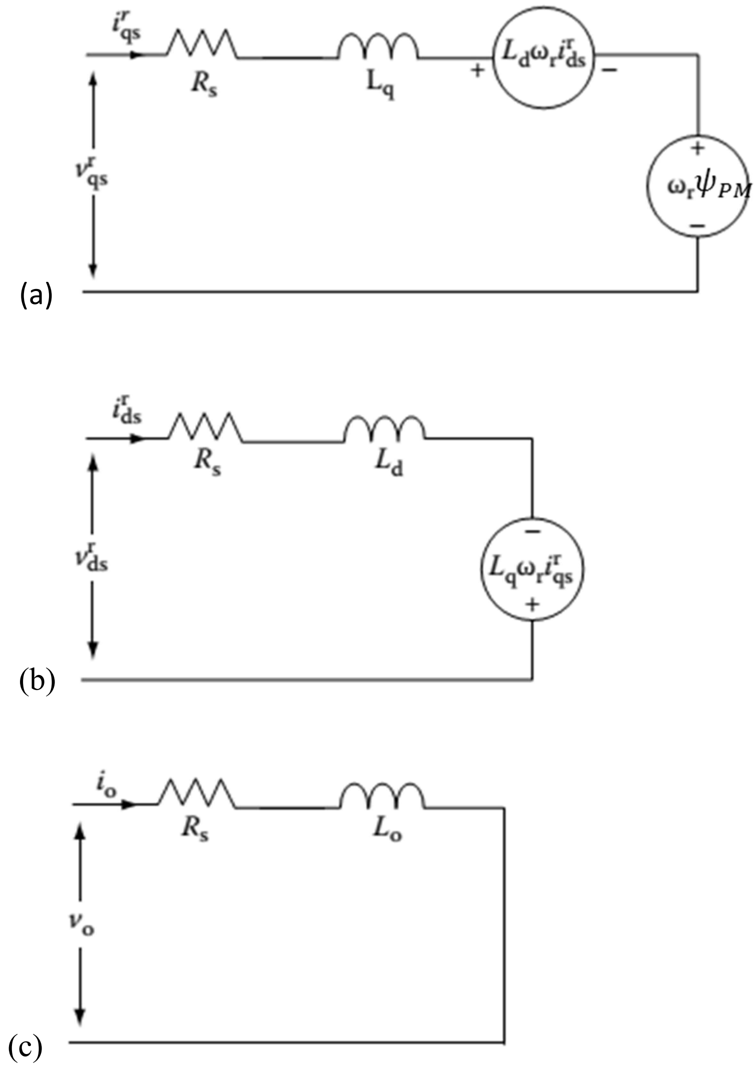
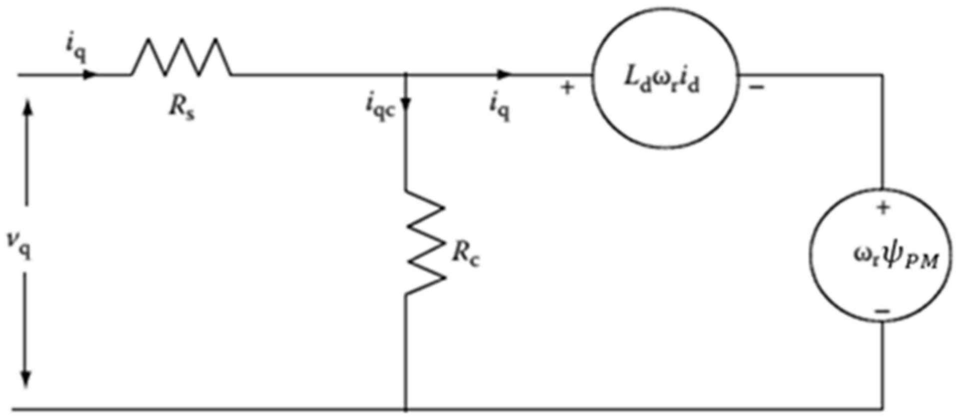
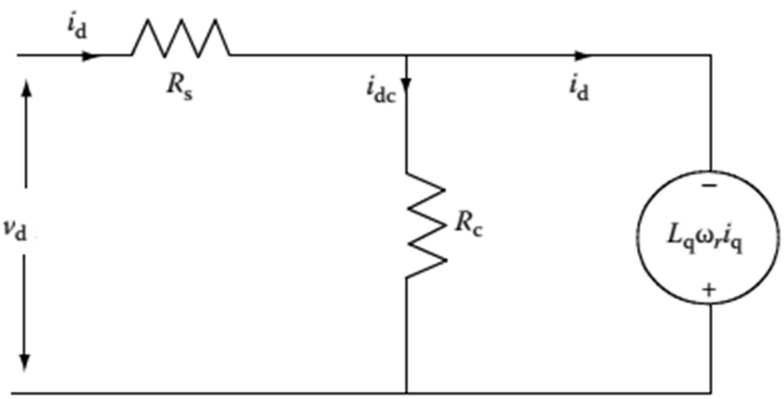


Figure.2.4, Dynamic equivalent circuits. (a) Dynamic stator q-axis. (b) Dynamic stator d-axis. (c) Zero sequence equivalent circuit of PMSM neglecting core losses [5].

The core losses can be represented by an equivalent resistance. The model useful in the study of optimal torque operation of the machine, and most significantly to determine the torque versus speed boundary for the optimal utilization and the safe operation of the machine. Only steady-state equivalent circuits with core-loss modeling are shown in Figure 2.5.



(a)



(b)

Figure 2.5, Steady-state equivalent circuits (a) Steady-state q-axis and (b) Steady-state d-axis equivalent circuit of PMSM with core losses [5].

2.5. Some common control schemes for PM motors

PMSM control methods broadly classified into scalar and vector based control. Scalar control schemes are based on the steady state relationships. Frequency and amplitude of a controlled space vector is considered. However, in vector control, magnitude and position of a controlled space vector is taken into consideration. These relations are valid even during transients which is essential for precise control of speed and torque.

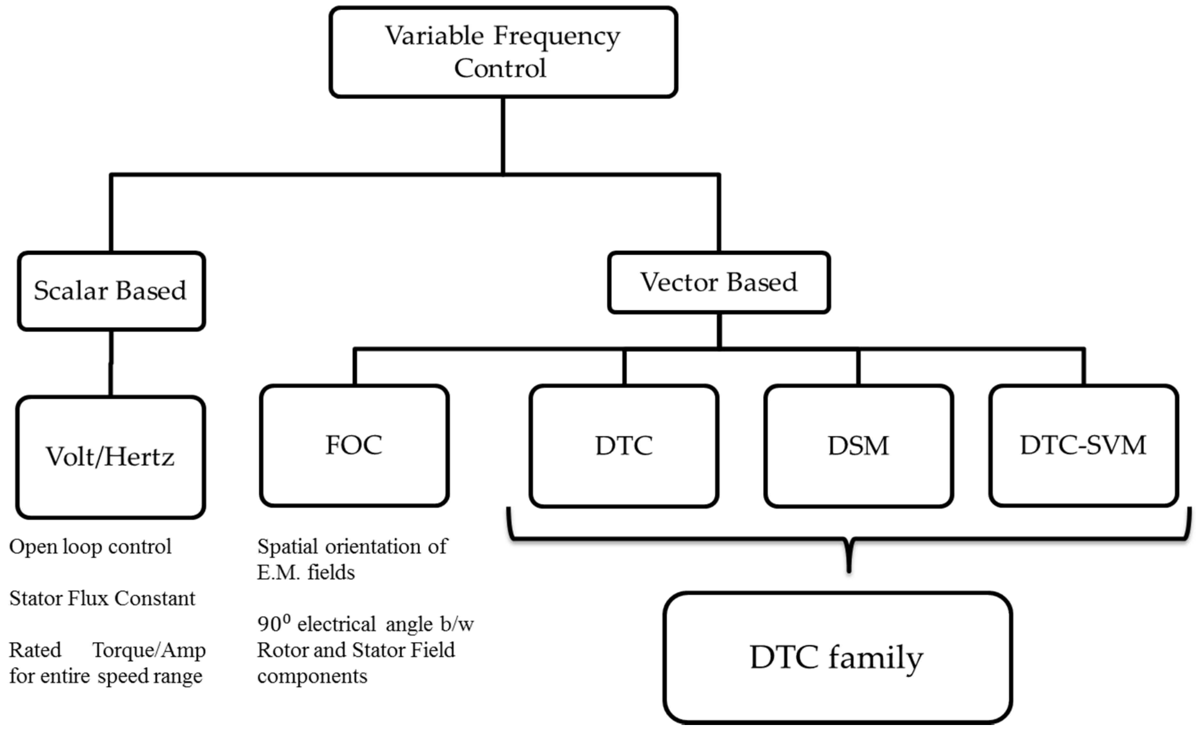


Figure 2.6, Some control schemes used for PMSM.

Summary

In this chapter, the mathematical model of the surface PMSM is derived in rotor-reference frame with respect to the rotor of PM motor. The dynamic stator d- and q-axis equivalent circuit of motor without core losses from the stator voltage equations and drawn which will be used for simulation. Different control strategies are also shown in a tree diagram. The DTC scheme and DTC-SVM scheme are to be discussed in following two chapters respectively.

Chapter III

Direct Torque Control (DTC) of Permanent Magnet Synchronous Motor (PMSM) Drives

- Introduction
- Control Scheme
- Flux and torque reference
- Flux Linkage Estimation
- Electromagnetic Torque Estimation
- Optimum switching voltage vector Selection
- Voltage Source Inverter

3.1.GENERAL INTRODUCTION

The control principle is to directly control the Torque and Stator flux of a drive by inverter voltage space vector selection through a lookup table[6][9][12]. The control algorithm is to pick a control signal that changes the quantity in problem if the actual value is other than the desired value, evenif the difference is small or big[15]-[18]. Torque and Flux signals are compared with their actual values and the error signals are compared with hysteresis controller, accordingly voltage vectors are selected. The benefits of the DTC scheme are low intricacy and only one parameter has to be used, i.e. the stator resistance[12]-[14]. All calculation for the DTC scheme is done in a stationary reference frame and hence no need to know the rotor angle. Only the rotor position at startup has to be known. It gives better performance with a speedy, dynamic response. But the disadvantages are high ripple levels of torque and stator flux linkages. The stator currents contain harmonics due to the crude selection of the voltage switches[15]-[17]. In case of a current source inverter fed motor, Optimal current switching vectors are selected while the optimal voltage switching vectors are selected for a voltage source inverter fed motor, [1],[7].

3.2.CONTROL SCHEME:

Here a direct torque controlled PMSM run by a voltage source inverter, VSI is presented. The electromagnetic torque stator and the flux linkages are controlled directly by applying optimal voltage switching vectors to the VSI. The principal goal is to select the voltage switching vectors which provide the fastest electromagnetic torque response. The six active switching vectors (V_1, V_2,.....V_6), and two zero switching vectors (V_7,V_0) is shown in fig. 3.1.

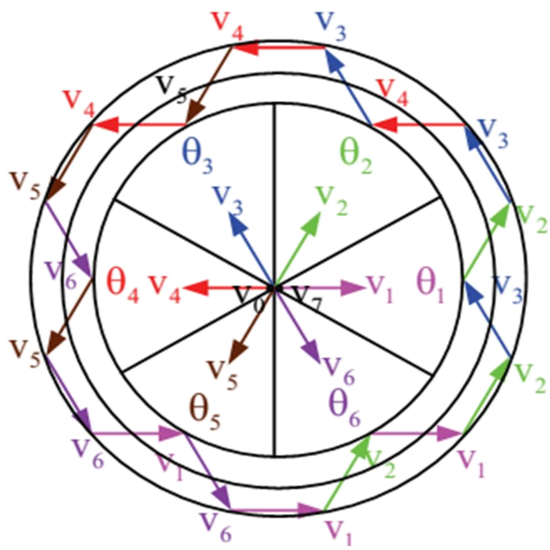


Figure 3.1, The switching vector and sector distribution to locate the flux position.

The six voltage switching vectors form six sectors. Each sector is within two active voltage vectors and span 60^0 electrical. For a six step inverter the minimal number of sector is six[6].

The DTC scheme of PMSM supplied by a VSI is given fig. 3.2.-

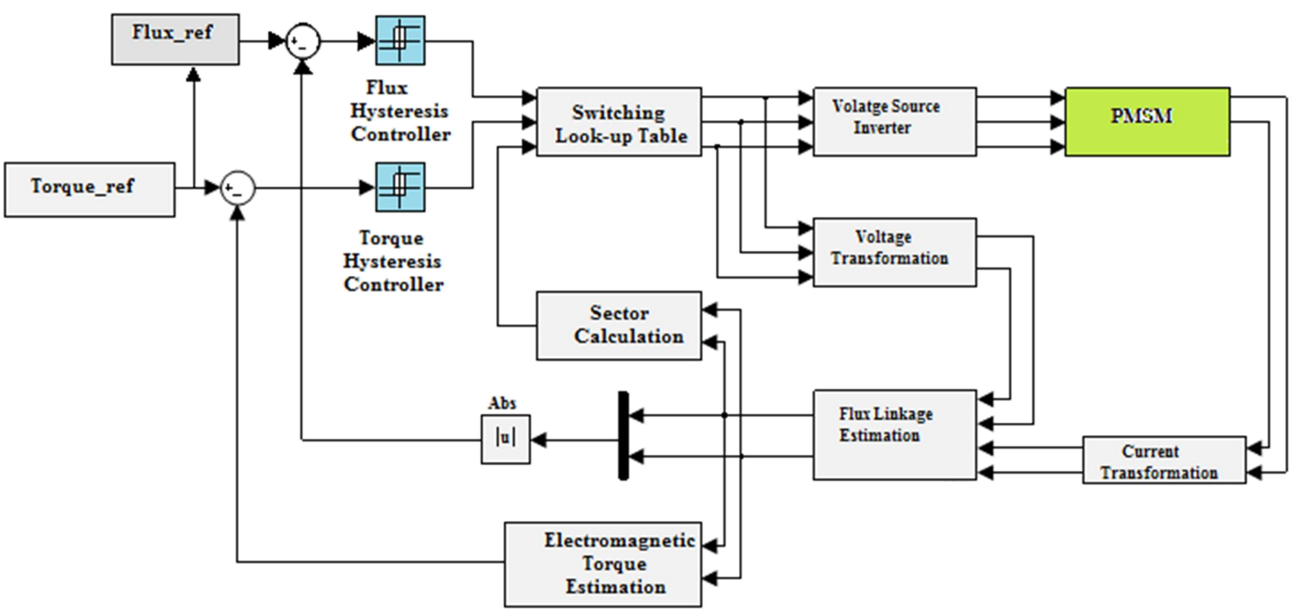


Figure 3.2, Control Scheme of the DTC PM Synchronous Motor supplied by a VSI.

The electromagnetic torque error and the stator flux-linkage error are inputs to the respective flux-linkage and torque hysteresis comparators. Both the flux-linkage and the electromagnetic torque comparators are two-level comparators. The discretized outputs of the hysteresis comparators ($d\psi_s$, dT_e) are inputs to the optimum voltage switching selection table. Sector, which determines the right vector is also an input to the switching table. for each sampling instant this procedure is repeated to get the wanted speed value.

3.3.STATOR FLUX LNKAGE AND TORQUE REFERENCE

The stator flux-linkage error is generated by comparing the estimated stator flux-linkage modulus to its reference value ($|\psi_{s\_ref}|$). However, this reference value is obtained from the reference value of the electromagnetic torque. So reference flux is obtained as a function of the reference electromagnetic torque, i.e. $\psi_{s\_ref}(|T_{e\_ref}|)$. The function $\psi_{s\_ref}(|T_{e\_ref}|)$ can be obtained by ensuring maximal torque per unit current up to base speed, and for above base speed, by ensuring maximal torque per unit flux[6].

The stator flux is given as:

$$|\psi_s| = \left(\psi_{sD}^2 + \psi_{sQ}^2\right)^{1/2} \qquad (3.1)$$

Where $\psi_s$is the stator flux vector, $\psi_{sD}$and ψ_{sQ} are the direct and quadrature axis stator flux linkages given by

$$\psi_{sD} = L_d i_{sD} + \psi_{PM}, \qquad (3.2)$$

$$\psi_{sQ} = L_q i_{sQ}. \qquad (3.3)$$

$i_{sD}$and i_{sQ} are the direct and quadrature stator currents, $L_d$and L_q are the direct and quadrature inductances and ψ_{PM} is the flux from the permanent magnets.

Using equations (3.1) -(3.3) the amplitude of stator flux can be given as

$$|\psi_s| = \left[(L_d i_{sD} + \psi_{PM})^2 + \left(L_q i_{sq}\right)^2\right]^{1/2}, \tag{3.4}$$

the electromagnetic torque is given by:

$$T_e = \frac{3}{2L_s}\left(\frac{P}{2}\right)\psi_{PM}|\psi_s|\sin\delta \tag{3.5}$$

where *P* is the number of poles. And also in terms of stator current can be given as

$$T_e = \frac{3}{2}\left(\frac{P}{2}\right)(\psi_{PM} i_{sQ} + (L_q - L_d) i_{sD} i_{sQ}), \tag{3.6}$$

in which, by considering $i_{sD} = 0$ the maximum torque per unit of current is obtained.

Then (3.6) is given as:

$$T_e = \frac{3}{4}(P)\psi_{PM} i_{sQ}. \tag{3.7}$$

Thus the current developed by the torque is

$$i_{sQ} = {}^{4T_e}\!/_{3P\psi_{PM}} \tag{3.8}$$

Substituting (3.8) in (3.4) the reference flux function obtain as:

$$\left|\psi_{s\_ref}\right| = \left[\left((L_d i_{sD} + \psi_{PM})^2 + {L_q}^2\left({}^{4T_{e\_ref}}\!/_{3P\psi_{PM}}\right)^2\right)\right]^{1/2}. \tag{3.10}$$

3.4.FLUX LINKAGE ESTIMATION

The stator flux linkages and electromagnetic torque, for a drive without a position sensor can be obtained by the voltage reconstructed from the d.c. link voltage (U_{dc}) and the switching states (S_A,S_B,S_C) of the six switching devices of a six-step voltage-source inverter shown in figure 3.3.

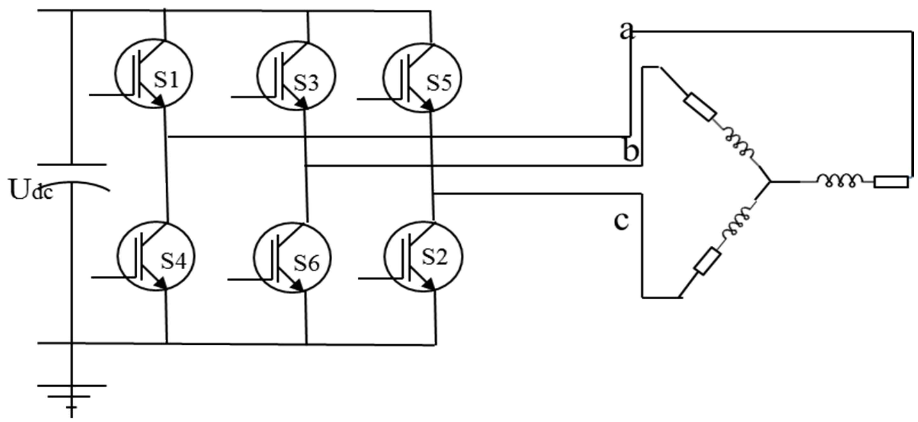


Figure 3.3, Voltage Source Inverter switches

The switching states (S_A for phase A, S_B for phase B, and S_C for stator C) are defined as in Table 3.1.

Table 3.1 Inverter Switching functions

S_A=l	S1 ON	S4 OFF
S_A=0	S1 OFF	S4 ON
S_B=l	S3 ON	S6 OFF
S_B=0	S3 OFF	S6 ON
S_C=l	S5 ON	S2 OFF
S_C=0	S5 OFF	S2 ON

The stator-voltage space vector in the stationary reference frame can be obtained by using the switching states and the d.c link voltage U_{dc} as:

$$V_s = \frac{2}{3} V_{dc}(S_A + aS_B + a^2 S_C) \tag{3.10}$$

The stator line to line voltage of PM synchronous machine can be expressed as

$$\begin{cases} V_{ab} = V_{dc}(S_A - S_B) \\ V_{bc} = V_{dc}(S_B - S_C) \\ V_{ca} = V_{dc}(S_C - S_A) \end{cases} \tag{3.11}$$

The stator phase voltages (line-to-neutral) can be obtained from the line-to-line voltages as

$$\begin{cases} V_{sA} = \frac{1}{3}(V_{ab} - V_{ca}) \\ V_{sB} = \frac{1}{3}(V_{bc} - V_{ab}) \\ V_{sC} = \frac{1}{3}(V_{ca} - V_{bc}) \end{cases} \tag{3.12}$$

Substituting equation (3.11) into (3.12), the stator phase voltages are given as

$$\begin{cases} V_{sA} = \frac{1}{3} V_{dc}(2S_A - S_B - S_C) \\ V_{sB} = \frac{1}{3} V_{dc}(-S_A + 2S_B - S_C) \\ V_{sC} = \frac{1}{3} V_{dc}(-S_A - S_B + 2S_C) \end{cases} \tag{3.13}$$

Now considering that $V_s = \frac{2}{3} V_{dc}(V_{sA} + aV_{sB} + a^2 V_{sC})$, we have the equation (3.10).

The stator flux space vector can be obtained by integration of the terminal voltage minus the stator ohmic drop.

$$\psi_s = \int (V_s - R_s i_s) dt \tag{3.14}$$

Direct and quadrature axis stator flux components in the stator reference frame are obtained as:

$$\psi_{sD} = \int (V_{sD} - R_s i_{sD}) dt \tag{3.15}$$

$$\psi_{sQ} = \int (V_{sQ} - R_s i_{sQ}) dt \tag{3.16}$$

By substituting equation (3.10) into equation (3.14), for a digital implementation the stator flux linkage space vector can be obtained as follows [6].The predicted stator flux linkage space vector for the kth sampled value of (in the stationary reference frame) can be expressed as:

$$\psi_s(k) = \psi_s(k-1) + \tfrac{2}{3} V_{dc} T [S_A(k-1) + aS_B(k-1) + a^2 S_C(k-1)]$$
$$-R_s T i_s(k-1) \tag{3.17}$$

where i_s is the stator current space vector and T is the sampling time constant.

The angle of the stator flux linkage space vector can be obtained from the direct and quadrature axis stator flux components (ψ_{sD} and ψ_{sQ}) as follows:

$$\phi_s = \tan^{-1}(\psi_{sQ} / \psi_{sD}) \tag{3.18}$$

3.5.ELECTROMAGNETIC TORQUE ESTIMATION

It is necessary to have a very accurate electromagnetic torque estimator for instantaneous torque control. The electromagnetic torque estimator in fig.(3.2) estimates the electromagnetic torque. Electromagnetic torque can be estimated by using the estimated flux linkages and monitored stator currents as:

$$T_e = \tfrac{3}{2}\left(\tfrac{P}{2}\right)(\psi_{sD} i_{sQ} - \psi_{sD} i_{sD}) \tag{3.19}$$

Using the stator currents in the rotor i.e., synchronous reference frame (i_{sd},i_{sq}),we can get the torque equation as

$$T_e = \tfrac{3}{2}\left(\tfrac{P}{2}\right)(\psi_{PM} i_{sq} + (L_q - L_d) i_{sd} i_{sq}). \tag{3.20}$$

3.6.OPTIMUM SWITCHING VOLTAGE VECTOR SELECTION

To get the fastest torque response is the primary objective of the switching table. The switching table presented by [13] (Table 2) takes into account the sector (θ_1, θ_2,...,θ_6) and the digital output levels of the flux and torque hysteresis controllers (dψ_s, dT_e). Fig. 3.1 shows that each sector covering 60 electrical degrees by the regions θ_1, θ_2,...,θ_6, where the sector 1 covers from –30° up to 30°, sector 2 from 30° to 90° and so on until 330° is reached. The plane is defined by the axes sD and sQ where sD is at zero degrees [6]. To determine the correct control commands one flux and one torque hysteresis comparators are used. The comparators evaluate the difference between requested values and estimated values and gives output in digital levels.

Flux Hysteresis Controller Output:

a) For $|\psi_s| < |\psi_{s\_ref}| - \Delta\psi_s$, $d\psi_s$=1 and the stator flux increases.
b) For $|\psi_s| > |\psi_{s\_ref}| + \Delta\psi_s$, $d\psi_s$=0 and the stator flux decreases.

Torque Hysteresis Controller Output:

a) For $|T_e| < |T_{e\_ref}| - \Delta T_e$, dT_e=1 and the electromagnetic torque increases
b) For $|T_e| > |T_{e\_ref}| + \Delta T_e$, dT_e=0 and the electromagnetic torque decreases

During the switching interval, each voltage vector is constant and to select the voltage vectors for controlling the amplitude of ψ_s, the voltage vectors adjacent to the sector where the flux is located, are to be selected to increase or decrease the amplitude as the closest vectors give the minimum switching frequency.

From equation (3.5), it can be seen that the torque is proporsional to the angle, δ, [12] not the slip frequency as in induction motors[9],[10]. Therefore, for controlling the amplitude of the stator flux linkage and for changing the torque quickly, zero voltage vectors are not used in PMSM[12]-[14]. In otherwords, ψ_s should always be in motion with respect the rotor flux linkage. The rotating direction of ψ_s is determined by the output of the torque controller. The switching table for controlling both the amplitude and rotating direction of ψ_s is as shown in Table 2 and is used for both direction of operations.

Table 3.2 Look Up Table for Appropriate Switching Vector Selection

Flux Error $d\psi_s$	Torque Error dT_e	S1	S2	S3	S4	S5	S6
1	1	V_2(110)	V_3(010)	V_4(011)	V_5(001)	V_6(100)	V_1(100)
	0	V_6(100)	V_1(100)	V_2(110)	V_3(010)	V_4(011)	V_5(001)
0	1	V_3(010)	V_4(011)	V_5(001)	V_6(100)	V_1(100)	V_2(110)
	0	V_5(001)	V_6(100)	V_1(100)	V_2(110)	V_3(010)	V_4(011)

3.7.VOLTAGE SOURCE INVERTER

The Switching Look-up Table block sends command to the Voltage Source Inverter (VSI) to generates the desired voltage. Here in DTC voltage synthesization is pretty easier since here no pulse width modulation is used; for the entire period of operation the out put devices stay in the same state. A simplified diagram of the VSI output stage is shown in figure 3.3.

The output signals from the Switching Look-up Table in Figure 3.2 are named as S_A, S_B and S_C. These are boolean variables indicating the switch state in the inverter output branches.

Let $S_i = 1$ when the upper switch is ON and the lower is OFF, and $S_i = 1$ when the lower switch is ON and the upper one is OFF. The voltage vectors generated by the inverter states S_A, S_B and S_C are given as:

$$V_1 = (1\ 0\ 0), V_2 = (1\ 1\ 0), V_3 = (0\ 1\ 0), V_4 = (0\ 1\ 1), V_5 = (0\ 0\ 1), V_6 = (1\ 0\ 1).$$

Summary

In this chapter the conventional Direct Torque Control scheme is described. The flux and torque estimators are presented. The switching conditions and the look up table is also described. The control principle with the block diagram and the flux control strategies are presented with figures. How the inverter voltages are synthesized with the DC link voltage is also explained.

Chapter IV

DTC using Space Vector Modulation (DTC-SVM) of Permanent Magnet Synchronous Motor (PMSM) Drives

- General Introduction Of Space Vector Pulse Width Modulation
- Space Vector Pulse Width Modulation Method
- Inverter States And Voltage Vectors Of A Two Level VSI
- Direct Torque Control By Using Space Vector Modulation (DTC-SVM) For The PMSM Electromagnetic Torque Estimation

4.1.INTRODUCTION OF SPACE VECTOR PULSE WIDTH MODULATION:

The space vector pulse width modulation (SVM) method is an advanced computation-intensive PWM method. It is possibly the best among all the PWM techniques for variable frequency drive applications. It is gaining its application in wide-range, because of its performance characteristics, in recent years.

In general, PWM methods only a half bridge of a three-phase bridge inverter is considered. All the half bridges operate independently when the load neutral is connected to the center tap of a DC supply. But, load neutral is usually isolated with a machine load which results in interaction among the phases. In other PWM methods, this interaction was not considered. The space vector pulse width modulation considers the combined effect of all the three-phase interactions and minimizes the harmonic content of the three phase isolated motor load.

4.2.SPACE VECTOR PULSE WIDTH MODULATION METHOD:

The Space-vector pulse width modulation (SVM) can be understood by using concept of revolving mmf in a three-phase machine [1]. The three phase sinusoidal voltages fed to three phase windings on the stator, which produce a revolving/ rotating mmf in the air gap or the space. This revolving mmf is an example of a space vector.

Let’s consider the mmf produced by a single phase winding, R as shown in Figure 4.1. A single-phase winding produces a pulsating magnetic field while a three-phase winding produces a revolving magnetic field. The pulsating magnetic field is given as:

$$F_R = K i_R \cos\theta_e \tag{4.1}$$

$$i_R = I_m \cos(\omega_e t) \tag{4.2}$$

$$F_R = KI_m \cos(\omega_e t) \cos\theta_e \tag{4.3}$$

$$F_R = \frac{F_{max}}{2}[\cos(\theta_e - \omega_e t) + \cos(\theta_e + \omega_e t) \tag{4.4}$$

$$F_R = F_R^+ + F_R^- \tag{4.5}$$

Where F_R is mmf along R-phase, i_R current flowing in R phase winding, I_m maximum value of currenti_R, θ_e is the angle between phase-R and mmf F_R, ω_e is the angular speed of the rotor, F_{max} is the maximum value of mmf, F_R^+ and F_R^- are anticlockwise and clockwise revolving mmfs.

The equation (4.5) represents a pulsating magnetic field. The pulsating magnetic field can be resolved into two revolving magnetic fields- one rotating anti clockwise and other rotating in clockwise direction.

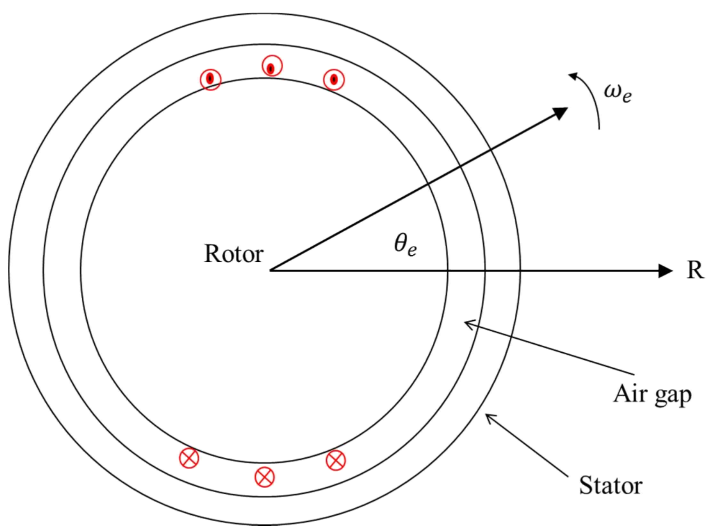


Figure 4.1 Mmf produced by a single phase, R winding.

Now consider a three-phase winding of the PMSM machine as shown in figure 4.2. Now when all the windings are excited by a three phase sinusoidal currents, what happens is analyzed as follows.

The sinusoidal currents in all the windings can be given as:

$$\begin{cases} i_R = I_m \cos(\omega_e t) \\ i_Y = I_m \cos(\omega_e t - 120^0) \\ i_B = I_m \cos(\omega_e t + 120^0) \end{cases} \tag{4.6}$$

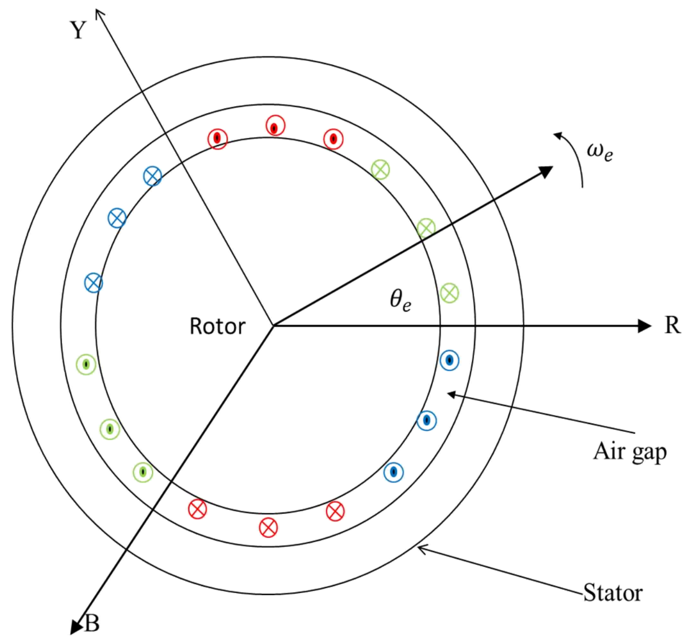


Figure 4.2, Mmf produced by three phase windings of PMSM.

Now the mmfs in all the three windings can be expressed as:

$$\begin{cases} F_R = K i_R \cos\theta_e \\ F_Y = K i_Y \cos(\theta_e - 120^0) \\ F_B = K i_B \cos(\theta_e - 240^0) \end{cases} \tag{4.7}$$

When we put equation (4.6) into (4.7), we have

$$\begin{cases} F_R = K I_m \cos(\omega_e t)\cos\theta_e \\ F_Y = K I_m \cos(\omega_e t - 120^0)\cos(\theta_e - 120^0) \\ F_B = K I_m \cos(\omega_e t + 120^0)\cos(\theta_e - 240^0) \end{cases} \tag{4.8}$$

$$\Rightarrow \begin{cases} F_R = \frac{F_{max}}{2}[\cos(\theta_e - \omega_e t) + \cos(\theta_e + \omega_e t) \\ F_Y = \frac{F_{max}}{2}[\cos(\theta_e - \omega_e t) + \cos(\theta_e + \omega_e t + 120^0) \\ F_B = \frac{F_{max}}{2}[\cos(\theta_e - \omega_e t) + \cos(\theta_e + \omega_e t + 240^0) \end{cases} \quad (4.9)$$

Total air gap flux produced by all the three windings is given as:

$$F_{ag} = (F_R^+ + F_Y^+ + F_B^+) + (F_R^- + F_Y^- + F_B^-) = 3F_R^+ \quad (4.10)$$

$$F_{ag} = \frac{3F_{max}}{2}\cos(\theta_e - \omega_e t) \quad (4.11)$$

The equation (4.11) is the expression of a revolving magnetic field. So we have a revolving magnetic field when the three stator windings of the rotor are excited by a three phase sinusoidal current.

We can transform the three phase system to an equivalent two phase system which produces the same revolving field by 3-phase to 2-phase transformation, i.e. Clarke's Transformation [1],[29].

The state space matrix-transformation of 3-phase currents in 2- phase currents can be given as:

$$\begin{bmatrix} i_\alpha \\ i_\beta \end{bmatrix} = \begin{bmatrix} \frac{3}{2} & 0 & 0 \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} i_R \\ i_Y \\ i_B \end{bmatrix} \quad (4.12)$$

Similarly the state space transformation of 3-phase voltages in 2-phase voltage is given below:

$$\begin{bmatrix} V_\alpha \\ V_\beta \end{bmatrix} = \begin{bmatrix} \frac{3}{2} & 0 & 0 \\ 0 & \frac{\sqrt{3}}{2} & \frac{\sqrt{3}}{2} \end{bmatrix} \begin{bmatrix} V_{RN} \\ V_{YN} \\ V_{BN} \end{bmatrix} \quad (4.13)$$

4.3.INVERTER STATES AND VOLTAGE VECTORS OF A TWO LEVEL VSI:

The voltage source inverter is as shown in figure 3.3. The six-step voltage source inverter has three legs with six power electronics devices or switches. In each leg there are two switches one in top and other in bottom. The top devices are ON when positive states appear and bottom devices are ON when negative states appear on them [1]. Accordingly there are 8 (2^3) combinations of the states. These 8 states give rise to 8 voltage vectors with six active (V_1, V_2,…V_3) and two zero (V_0 and V_7) voltage vectors as shown in figure 4.3.

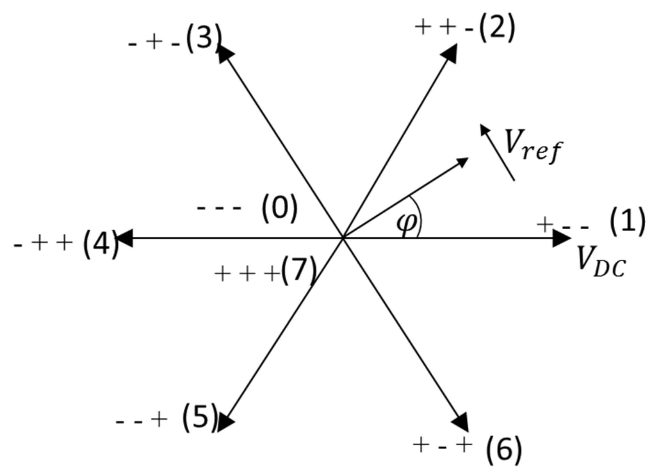


Figure 4.3 Inverter states and voltage vectors of a two level VSI.

In zero voltage-vectors, all the three phases are shorted and no power conversion is taking place from DC to AC side. For V_0 (- - -) voltage state, all the bottom devices are ON while for V_7 (+ + +) all the top devices are ON making the ac side a short circuit [28]-[29]. For active voltage vectors not all the top or bottom devices are ON (or OFF) simultaneously but either one or two of the top devices with other one or two bottom devices are on for a particular state causing power conversion from DC to AC side.. For example, in case of V_1 (+ - -) the top device of R-phase and the bottom devices of Y- and B-phase are ON and corresponding bottom/ top devices are OFF. Again for V_2 (+ + -) the top devices of R- and Y-phase are ON and bottom device of R-phase is ON and the corresponding bottom/ top devices are OFF. Similarly, for all the other four states as stated in figure 4.3. All the active voltages have same magnitude V_{DC} but the angle between them changes as shown in figure 4.3, but zero vectors have no magnitude [28].

From the inverter we want to synthesize three-phase sinusoidal voltage to the PM synchronous motor. On the space vector plane three phase sinusoidal voltages mean revolving/ rotating voltage vector V_{ref} of constant magnitude of ${}^{3}/_{2}\, V_m$, where $V_m$is the peak phase voltage and it rotates at an angular velocity of fundamental frequency[28][29].

Let's consider the space voltage vector V_{ref} rotates on the state space and is at some arbitrary position on sector S1 which is an angle α from the R-phase axis which is along the voltage vector,V_1we sample this rotating reference voltage space vector V_{ref} with a high sampling frequency or a low sampling period[28]-[29]. If the sampling frequency is very high then the switching frequency for the inverter will be very high which gives rise to much more switching losses[28]-[33]. So there is a limit to the sampling frequency which depends on the inverter. Very high sampling frequency ensures a very good and continuous sinusoidal voltage output from the inverter. This frequency is equivalent to the carrier frequency where the frequency of the sine is proportional to the continuously rotating reference space vector and the period proportional to the triangular period. During a sampling period the magnitude of the reference space voltage vector is considered constant. For the reference vector in sector S1 it is synthesized by applying the voltage vectors V_1, V_2 and the zero vectors such that the volt-sec is balanced[30]-[32].

Volt-sec balance

Suppose we have a 10V DC source and we want 5V from this source. To get the average value of 5V, the 10V DC signal is chopped using a switch. When the switch is closed, whole 10V is applied to the output and when switch is opened, no voltage appears. So to get the average value of 5V, the volt-sec balance is given as

$$5V \times T_s = 10V \times T_{on} + 0 \times T_{off}$$

$$\Longrightarrow 5V = 10 \times \frac{T_{on}}{T_s}$$

This volt-sec balance is also applicable to the inverter operation or for the voltage space vector synthesis[30]-[32]. Consider the figure 4.4 in which the volt-sec balance is taken along α and β-axes.

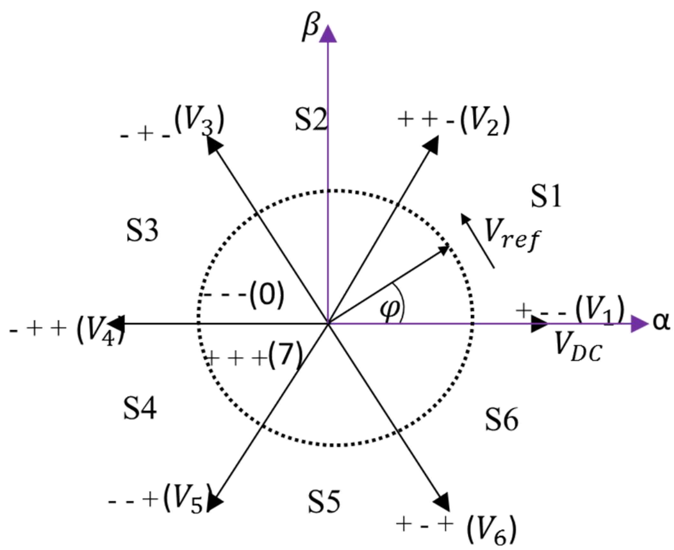


Figure 4.4 Space states of a two level VSI synthesizing the reference voltage vector V_{ref}.

The component of V_{ref} along α and β-axes multiplied by the sampling period i.e., voltage of the reference voltage along α and β-axis should be equal to the volt-sec of V_1, V_2 and the zero vectors along the α and β-axis.

For sector S1 is is explained here and it is applicable to all the sectors with voltage remaining same only the angle is changed. In sector S1 to synthesize reference space voltage vector V_1,is switched for a period T_1, V_2 is switched for a period of T_2 and for the remaining period the zero vectors are switched. The switching periods T_1 T_2 and T_0 can be calculated or determined for a sampling period as follows.

$$T_0 = T_s - (T_1 + T_2) \tag{4.14}$$

The volt-second balance equation along α-axis is

$$|V_{ref}| \cos\varphi \,.\, T_s = |V_1| . T_1 + |V_2| \cos 60^0 \,.\, T_2 \tag{4.15}$$

And the volt-second balance equation along β-axis is given as

$$|V_{ref}| \sin\varphi \,.\, T_s = 0. T_1 + |V_2| \sin 60^0 \,.\, T_2 \tag{4.16}$$

Solving the equations (4.15) and (4.16) we get the time periods as follows:

$$T_1 = \frac{2}{\sqrt{3}} . \frac{T_s |V_{ref}|}{V_{dc}} \sin(60^0 - \alpha) , \tag{4.17}$$

$$T_2 = \frac{2}{\sqrt{3}} . \frac{T_s |V_{ref}|}{V_{dc}} \sin\alpha \tag{4.18}$$

and, $T_0 = T_s - (T_1 + T_2)$ (4.19)

The pattern of switching of these voltage vectors should be such that minimum switching loss occurs. So, for minimum switching loss we should ensure minimum switching of the inverters. The time period for zero voltage switching is divided into two equal intervals as

$$T_{01} = T_{02} = {}^{T_0}/_{2}.$$

The switching pattern for two consecutive sampling periods is given in figure 4.5 witch the inverter output voltages.

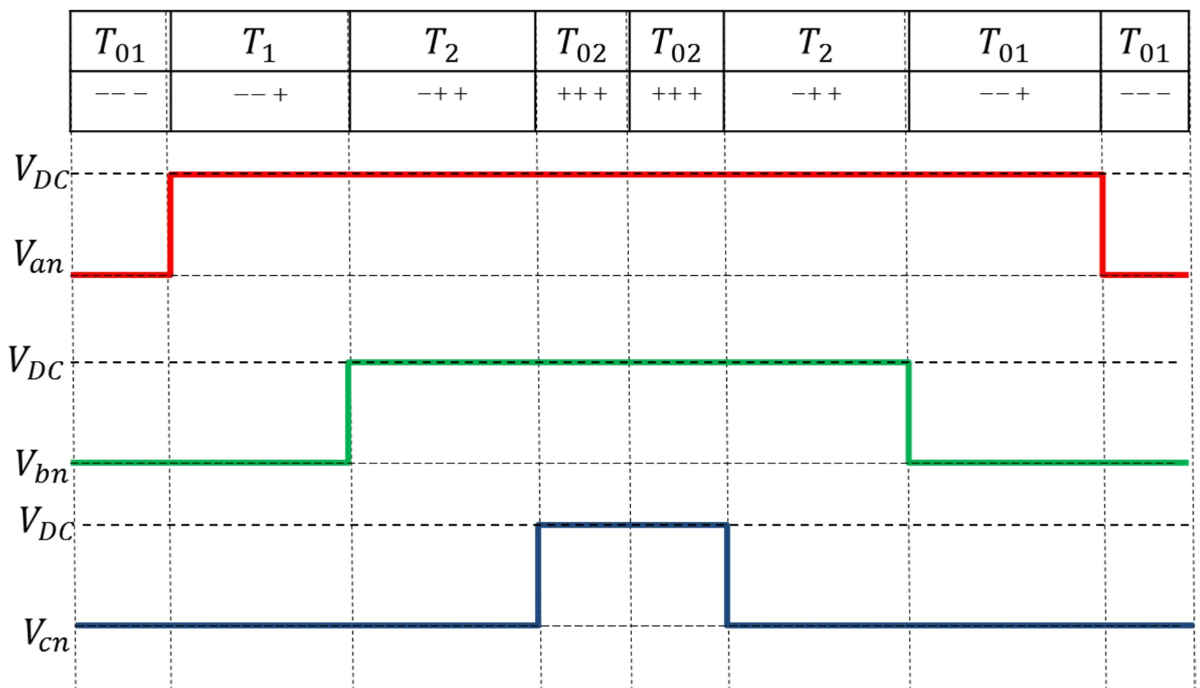


Figure 4.5 Switching patterns of the inverter states for minimum switching and the output phase voltages [28]-[29].

4.4.DIRECT TORQUE CONTROL BY USING SPACE VECTOR MODULATION (DTC-SVM) FOR THE PMSM:

The objective of DTC scheme is to maintain the torque and flux linkage close to their reference values by appropriately selecting the inverter output voltages. The SVM scheme used here is to have the desired torque and flux linkage values by selecting the inverter states. The schematic block diagram the DTC scheme by using SVM technology for the PMSM is as shown in figure 4.6 [22]-[27]. As shown in figure the hysteresis controllers and the switching look up table are replaced by the PI controllers and Discrete Space Vector PWM block respectively [22].

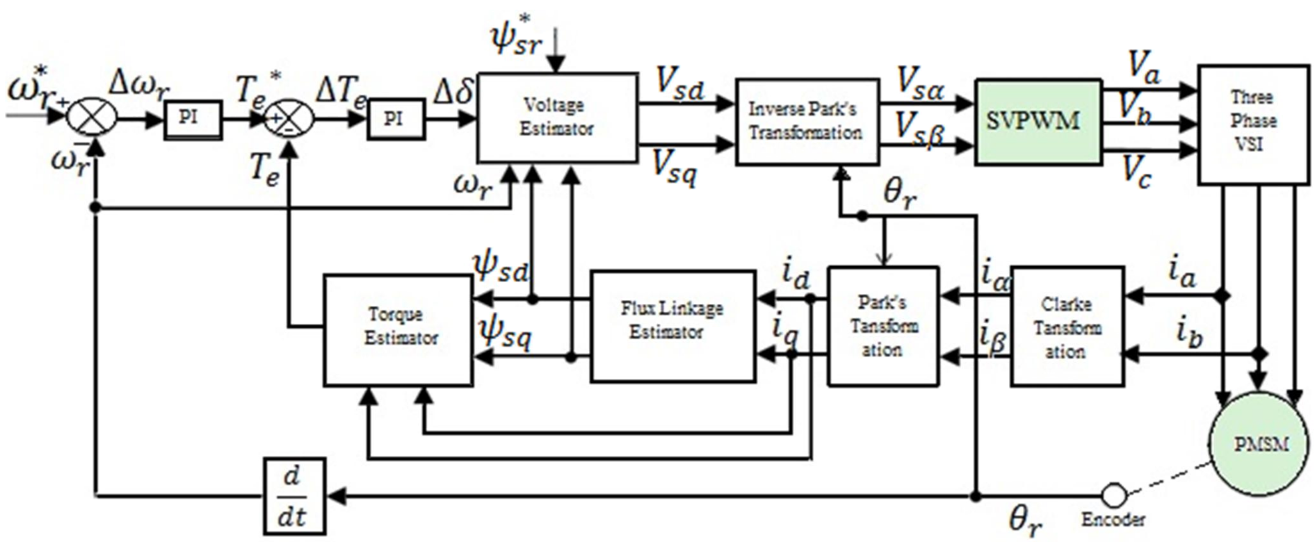


Figure 4.6 Schematic Block diagram of DTC-SVM scheme of the PMSM drive system.

The stator linkage flux and electromagnetic torque are estimated using the stator currents and the stator inductances. The stator currents are transformed from abc-frame to $\alpha\beta$-frame and then to dq-frame. These dq-currents are used for estimation of the actual flux linkages and the electromagnetic torque.

The electromagnetic torque for the PMSM drive is given as:

$$T_e = \frac{3}{2L_s}\left(\frac{P}{2}\right)\psi_{PM}|\psi_s|\sin\delta \tag{4.20}$$

From the above equation, the electromagnetic torque is directly proportional to the sine of the angle between the stator and rotor flux linkages, δ [22]-[24]. By changing the angle δ quickly we can change the torque and flux at a faster rate. The reference torque which is to be tracked and the actual torque are compared and the sampled error is given to PI controller to get the change flux angle, $\Delta\delta$. This $\Delta\delta$ is added to the actual angle to get the new position of the flux. The reference stator flux linkage, the error signal $\Delta\delta$, the estimated flux linkages and the dq-currents are fed to a voltage predictor which estimates reference voltage vector which is used for the space vector modulation as given in equations (4.17)-(4.19).

The reference space vector can be estimated using the error angle,$\Delta\delta$, the stator resistance R_s, actual stator flux ψ_s and current as:

$$V_{s\alpha\_ref} = \frac{\psi_{s\_ref}\cos(\delta+\Delta\delta)-\psi_s\cos\delta}{T_s} + R_s i_{s\alpha} \tag{4.21}$$

$$V_{s\beta\_ref} = \frac{\psi_{s\_ref}\sin(\delta+\Delta\delta)-\psi_s\sin\delta}{T_s} + R_s i_{s\beta} \tag{4.23}$$

$$V_{s\_ref} = \sqrt{{V_{s\alpha\_ref}}^2 + {V_{s\beta\_ref}}^2}, \varphi_{Vref} = \tan^{-1}\left(\frac{V_{s\beta\_ref}}{V_{s\alpha\_ref}}\right) \tag{4.24}$$

where T_s is the sampling period.

For operating in constant flux region, the reference stator flux linkage ψ_{s_ref} has magnitude equal to the flux magnitude of the permanent magnets ψ_{PM}. The reference space voltage vector is used in the SVPWM module which generates the inverter voltages which drives the PMSM drives to the closest of the desired values of torque and flux.

Summery

In chapter IV the basic space vector modulation technique is explained. The inverter states of a two level VSI is described. How the sinusoidal three-phase supply is equivalent to a revolving space vector is also described. The application SVM technique with the direct torque control scheme is presented. The DTC-SVM of PMSM is explained with the control block diagram. The implementation of DTC-SVM scheme with the resilts is presented in Chapter V.

Chapter V

Simulation Results and Discussion

- Result During Steady State For Conventional DTC Scheme Control Scheme
- Result During Steady State For DTC-SVM Scheme Control Scheme
- Result During Transient Conditions For Conventional DTC Scheme Control Scheme
- Result During Transient Conditions For DTC-SVM Scheme Control Scheme

SIMULATION RESULTS AND DISCUSSION

In chapters III and IV, the mathematical modelling of classical DTC and SVM-DTC schemes are described. The above schemes are implemented by developing a 2.5kW PMSM. The system parameters for the surface PMSM is given in Table 5.1. In the MATLAB/Simulink model designed for direct torque and flux control, two loops are integrated: one is the stator flux loop and other is the electromagnetic or the speed loop. Here the speed loop can be called as the outer loop to the electromagnetic torque loop. The main aim of this project work is to analyse and compare the performance of the classical-DTC and the SVM-DTC for the control of the stator flux and the electromagnetic torque and to select the control-scheme which gives ripple free torque response with lower harmonics in the stator currents. For classical DTC hysteresis controllers of bandwidth1Nm and 0.02Wb for electromagnetic torque and stator flux respectively

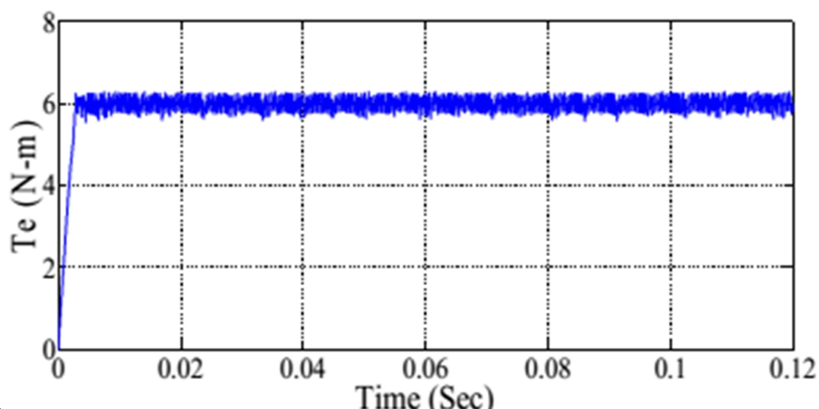
Table 5.1: PMSM MODEL PARAMETERS

Rated Power and Voltage	2.5kW,220V
Rated Current	3A
Permanent Magnet flux linkage ψ_{PM}	0.272Wb
Stator Inductances $L_d = L_q = L_s$	52.5mH
No. of Poles, P	8
Stator winding Resistance, R_s	1.96Ω
Moment of Inertia, J	0.000179kg.m^2
Friction Co-efficient, B	0.05Nm/rad/s
DC-link Voltage, V_{dc}	300V
Nominal Speed	3000rpm
Sampling frequency for DTC	100KHz
Sampling Frequency for DTC-SVM	10KHz

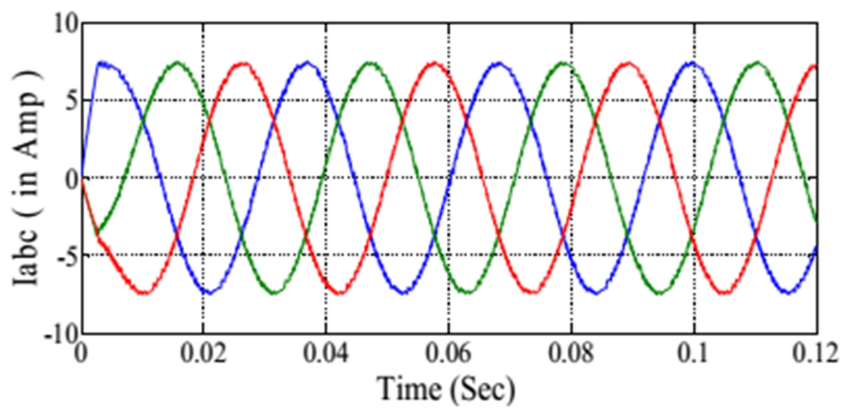
are used. The switching Table 3.2 is implemented using logic diagrams. For SVM-DTC implantation, a SVPWM module is developed using the equations (4.17)-(4.19).

5.1 DURING STEADY STATE FOR CONVENTIONAL DTC SCHEME:

For steady state analysis, a constant load of 6N-m with a speed reference of 200rad/s is applied at time $t \geq 0$sec. the simulation results are shown in figure 5.1. Figure 5.1(a) shows the electromagnetic torque response where the steady state value is achieved in less than 5msec but the ripple content is much high. The torque ripple in this case is around 0.6 N-m which is very close to the set hysteresis band. Figure 5.1(b) shows the actual stator current response.



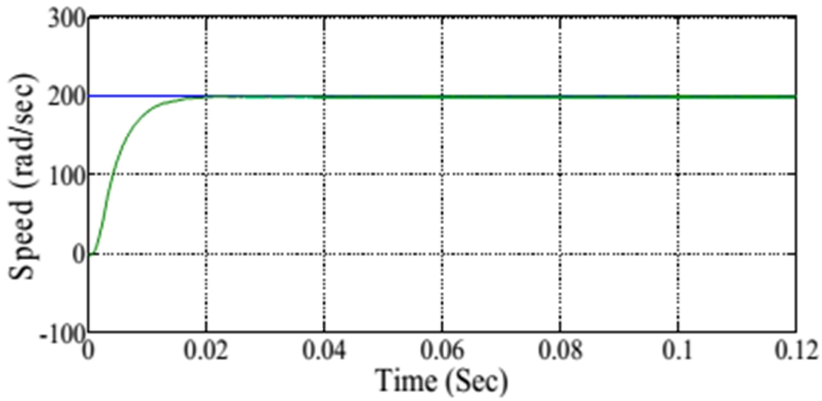
(a)



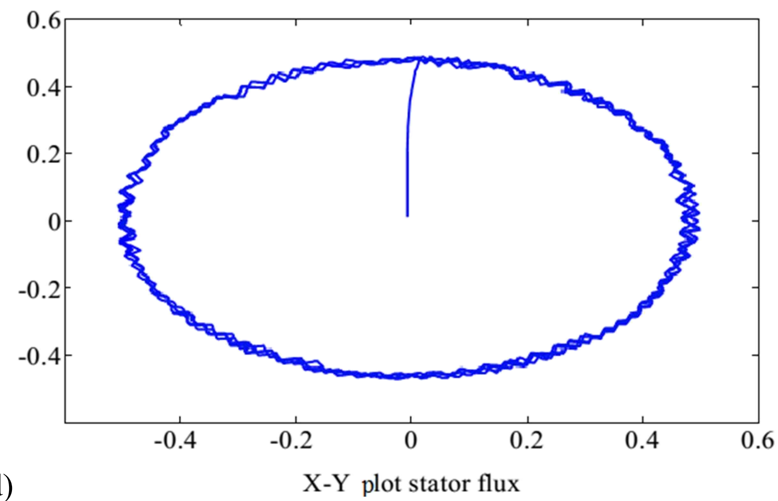
(b)

Figure 5.1 (a) Electromagnetic torque developed, (b) Actual stator current of the PMSM drive system for conventional DTC scheme during steady state conditions.

The speed response is shown in figure 5.1(c), which shows that the reference speed is achieved below 20msec. Figure 5.1(d) shows the X-Y plot stator flux variation containing large amount of ripples as the hysteresis band is fixed.



(c)



(d)

Figure 5.1 (c) Rotor Speed response, (d) X-Y plot of stator flux linkage of the PMSM drive system for conventional DTC scheme during steady state

5.2 RESULT DURING STEADY STATE FOR DTC-SVM SCHEME:

In this case all the conditions for DTC is same and the SVPWM modulator is applied to get the switching operation of VSI. From the figures 5.2(a) and 5.2(d), it is clearly observed that the electromagnetic torque and stator current responses are very smooth drastically reducing the